New Resistorless Current-Mode Quadrature Oscillators Using 2 CCCDTAs and Grounded Capacitors

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Abstract. *The current-mode quadrature oscillators using 2 current controlled current differencing transconductance amplifiers (CCCDTAs) and 2 grounded capacitors are presented. The proposed oscillators can provide 2 sinusoidal output currents with 90^o phase difference. The oscillation condition and oscillation frequency can be electronically/independently controlled by adjusting the bias current of the CCCDTA. High output impedances of the configuration enable the circuit to drive the external load without additional current buffers. The use of only grounded capacitors is ideal for integration. The PSpice simulation results are depicted. The given results agree well with the theoretical anticipation.*

Keywords

Sinusoidal oscillator, CCCDTA, current-mode.

1. Introduction

In the field of electric and electronic engineering, oscillators play an important role and have been widely applied in various aspects such as communications systems, instrumentation, measurement and signal processing, etc. The concept of oscillator design has been mainly on the requirement of multiple sinusoids which are 90° phase shifted, called quadrature signal, for easy implementation with other circuits for example in the design of SSB modulator [1], etc. From the past, there have been attempts to synthesis the sine wave oscillator in both forms of current and voltage mode. In the last decade, there has been a necessity to reduce voltage consumption in the circuit to support the wireless devices that run on compact batteries. Such requirement calls for the development of current-mode circuit designs due to their potential advantages such as inherently wide bandwidth, higher slew-rate, greater linearity, wider dynamic range, simple circuitry and low power consumption [2-5].

In 2003, a new active building block, namely current differencing transconductance amplifier (CDTA) [6] is presented as an alternative to the current-mode circuit. CDTA seems to be a versatile component in the realization of analog signal processing circuits; especially analogue frequency filters [7-8]. However the parasitic resistances at the input ports cannot be electronically adjusted. So in some circuits design, there is a requirement for additional resistors to be associated with or multiple CDTA merged together which is not suitable to create an integrator circuit. Later, the modified version of CDTA wherein the parasitic resistances at current input ports can be electronically controlled by bias current has been proposed. This CDTA is called current controlled current differencing transconductance amplifier (CCCDTA) [9].

From literature survey, it is found that several implementations of oscillator employing CDTAs or CCCDTAs have been reported [10-22]. Unfortunately, these reported circuits suffer from one or more of following weaknesses: use more than two CDTAs or CCCDTAs and excessive use of the passive elements which is not convenient to further fabricate in IC, some reported circuits use multiple-output CDTA or CCCDTA. Consequently, the circuits become more complicated. The proposed quadrature oscillators (QO) using CDTA, CCCDTA and OTA are compared with previously published QOs of [10-36] and the results are shown in Tab. 1.

The aim of this paper is to introduce the high output impedance current-mode quadrature oscillators, based on CCCDTAs. The oscillation condition and oscillation frequency can be independently adjusted by electronic method. The circuit constructions consist of 2 CCCDTAs and 2 grounded capacitors. The PSPICE simulation results are also shown, which are in correspondence with the theoretical analysis.

2. Theory and Principle

2.1 Basic Concept of CCCDTA

The principle of the CCCDTA was published in 2006 by W. Jaikla and S. Siripruchyanun [9]. It was modified

Ref	**Active element**	**Number of active element**	**Non-interactive control for CO and FO**	**Grounded C only**	**Number of R+C**	**Electronic tune of CO and FO**	**Current-mode QO output**
[10]	CDTA	3	Yes	Yes	0+3	Yes	Yes
[11]	CDTA	3	Yes	Yes	0+3	Yes	Yes
[12]	CDTA	2	Yes	No	4+2	No	Yes
[13]	CDTA	2	Yes	Yes	1+2	No	Yes
[14]	CDTA	2	Yes	No	4+2	No	Yes
[15]	CDTA	3	Yes	Yes	0+2	Yes	Yes
[16]	CDTA	4	Yes	Yes	0+2	Yes	Yes
[17]	CDTA	3	Yes	Yes	0+2	Yes	Yes
[18]	CDTA	1	Yes	No	2+2	No	No
[19]	CDTA	1	No	No	1+2	No	Yes
[20]	MO-CCCDTA	1	Yes	Yes	0+2	Yes	No
[21]	CCCDTA	2	Yes	Yes	0+2	Yes	No
[22]	MO-CCCDTA	1	Yes	Yes	0+2	Yes	Yes
[23]	MO-CCCDTA	1	Yes	No	2+2	Yes	No
[24]	OTA	3 (Fig. 5a)	Yes	Yes	0+2	Yes	No
		4 (Fig. 5b)	Yes	Yes	0+2	Yes	No
		6 (Fig. 6)	Yes	Yes	0+2	Yes	No
[25]	OTA	3	Yes	Yes	0+2	Yes	No
[26]	OTA	4	Yes	Ye	1 (R_N)+2	Yes	No
[27]	CCII, OTA	4	Yes	No	0+2	Yes	No
[28]	OTA	2	Yes	Yes	1+2	Yes	No
[29]	OTA	2 (Fig. 2a)	Yes	No	0+3	Yes	No
		3 (Fig. 2b)	Yes	Yes	0+2	Yes	No
		4 (Fig. 2c-d)	Yes	Yes	0+2	Yes	No
		4 (Fig. 2e)	Yes	No	0+4	Yes	No
[30]	OTA	3 (Fig. 1f)	Yes	Yes	0+2	Yes	No
		4 (Fig. 1g-h)	Yes	Yes	0+2	Yes	No
		5 (Fig. 1d-e, i)	Yes	Yes	0+2	Yes	No
		6 (Fig. 1a-c)	Yes	Yes	0+2	Yes	No
[31]	OTA	2	Yes	No	1+2	Yes	No
[32]	OTA	2	Yes	Yes	1+2	Yes	No
[33]	OTA	3	Yes	Yes	0+2	Yes	Yes
[34]	OTA	2 (Fig. 3, 10)	Yes	No	1+2	Yes	No
		2 (Fig. 8)	Yes	No	3+2	Yes	No
[35]	OTA	3	Yes	Yes	0+2	Yes	No
[36]	OTA	2	No	No	0+2	Yes	No
Proposed QOs	CCCDTA	2	Yes	Yes	0+2	Yes	Yes

CO: condition of oscillation

FO: frequency of oscillation

R_N: Nonlinear resistor

Tab. 1. Comparison between various QOs using CDTA and CCCDTA.

from the first generation CDTA [6]. The schematic symbol and the ideal behavioral model of the CCCDTA are shown in Fig. 1(a) and (b). It has finite input resistances: R_p and R_n at the p and n input ports, respectively. These intrinsic resistances are equal and can be controlled by the bias current I_{B1}. The difference of the i_p and i_n input currents flows from port z. The voltage v_z on z terminal is transferred into current using transconductance g_m, which flows into output terminal x. The g_m is tuned by I_{B2}. In general, CCCDTA can contain an arbitrary number of x terminals, providing currents I_x of both directions. The characteristics of the ideal CCCDTA are represented by the following hybrid matrix:

$$\begin{bmatrix} V_p \\ V_n \\ I_z \\ I_x \end{bmatrix} = \begin{bmatrix} R_p & 0 & 0 & 0 \\ 0 & R_n & 0 & 0 \\ 1 & -1 & 0 & 0 \\ 0 & 0 & 0 & \pm g_m \end{bmatrix} \begin{bmatrix} I_p \\ I_n \\ V_x \\ V_z \end{bmatrix} \tag{1}$$

If the CCCDTA is realized using BJT technology, R_p, R_n and g_m can be respectively written as

$$R_p = R_n = \frac{V_T}{2I_{B1}}, \tag{2}$$

and

$$g_m = \frac{I_{B2}}{2V_T}. \tag{3}$$

V_T is the thermal voltage. I_{B1} and I_{B2} are the bias current used to control the parasitic resistances and transconductance, respectively.

2.2 General Structure of Quadrature Oscillator

The oscillator is designed by cascading the gain controllable lossy integrator and the inverting lossless integrator as systematically shown in Fig. 2. From block diagram in Fig. 2, the characteristic equation is written as

$$s^2ab + sb(1-k) + k = 0. \tag{4}$$

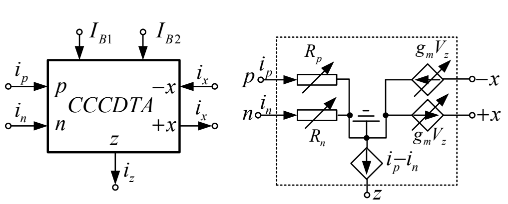


Fig. 1. CCCDTA (a) Symbol, (b) Equivalent circuit.

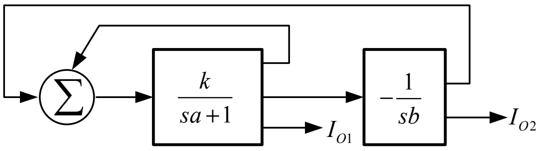


Fig. 2. Implementation block diagram for the quadrature oscillator.

From (4), the oscillation condition (OC) and oscillation frequency (ω_{osc}) can be written as

$$1 = k, \tag{5}$$

and

$$\omega_{osc} = \sqrt{\frac{k}{ab}}. \tag{6}$$

Considering (5) and (6), the oscillation condition can be controlled by the gain k, while the oscillation frequency can be changed by the natural frequency a, b or the gain k.

2.3 Proposed Current-Mode Quadrature Oscillators

The proposed quadrature oscillators are based on cascading of gain controllable lossy integrator and the inverting lossless integrator as shown in the last section. From block diagram in Fig. 2, the realization of proposed oscillators is achieved in Fig. 3(a) to (c). It is seen that the proposed circuits are resistorless and using only 2 grounded capacitors. Therefore, they are suitable IC implementation. Routine analysis, the characteristic equation of circuits in Fig. 3(a) and (b) is written as

$$s^2\frac{C_1C_2R_{n1}}{2g_{m2}} + s\frac{C_1}{g_{m2}}\left(1 - \frac{g_{m1}R_{n1}}{2}\right) + \frac{g_{m1}R_{n1}}{2} = 0. \tag{7}$$

while the characteristic equation of the circuit in Fig. (c) is shown as following:

$$s^2\frac{C_1C_2R_{p1}}{g_{m2}} + s\frac{C_2}{g_{m2}}\left(1 - \frac{g_{m1}R_{n1}}{2}\right) + \frac{g_{m1}R_{n1}}{2} = 0. \tag{8}$$

According to (5), the oscillation condition of all proposed oscillators is as follows:

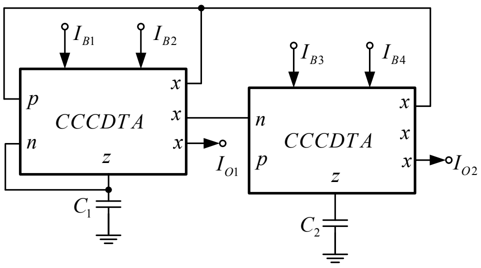
$$\text{OC: } 1 = \frac{g_{m1}R_{n1}}{2}. \tag{9}$$

According to (6), the oscillation frequency of proposed oscillators in Fig. 3(a) and (b) are as follows:

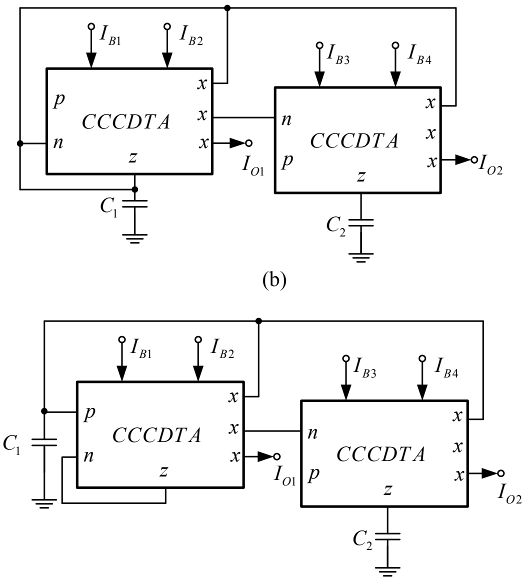
$$\omega_{osc} = \sqrt{\frac{g_{m1}g_{m2}}{C_1C_2}}. \tag{10}$$

while the oscillation frequency of circuit in Fig. 3(c) is written as

$$\omega_{osc} = \sqrt{\frac{g_{m1}g_{m2}}{2C_1C_2}}. \tag{11}$$



(a)



(b)

(c)

Fig. 3. Proposed quadrature oscillators.

Equation (11) is invalid if R_p and R_n are mismatch. Taking into account the mismatch of R_p and R_n, the oscillation frequency of the circuit in Fig 3(c) is written as

$$\omega_{osc} = \sqrt{\frac{g_{m1}g_{m2}R_{n1}}{C_1C_2R_{p1}}} . \tag{12}$$

Substituting the parasitic resistances and transconductance as shown in (2) and (3) into (9) to (11), the oscillation condition for all oscillators becomes

$$\text{OC: } 8I_{B1} = I_{B2} , \tag{13}$$

and the oscillation frequency of quadrature oscillator in Figs 3(a) and (b) is written as

$$\omega_{osc} = \frac{1}{2V_T}\sqrt{\frac{I_{B2}I_{B4}}{C_1C_2}} . \tag{14}$$

The oscillation frequency of circuit in Fig. 3(c) becomes

$$\omega_{osc} = \frac{1}{2V_T}\sqrt{\frac{I_{B2}I_{B4}}{2C_1C_2}} . \tag{15}$$

From (13) to (15), it can be seen that the oscillation condition can be adjusted electronically/independently from the oscillation frequency by varying I_{B1} while the oscillation frequency can be electronically adjusted by I_{B4}. From circuits in Fig. 3, the relationship between the explicit-current-outputs can be found as

$$\frac{I_{O2}(s)}{I_{O1}(s)} = -\frac{g_{m2}}{sC_2} . \tag{16}$$

For sinusoidal steady state, equation(15) becomes

$$\frac{I_{O2}(\omega_{osc})}{I_{O1}(\omega_{osc})} = \frac{g_{m2}}{\omega_{osc}C_2}e^{90} . \tag{17}$$

It is evident from (17) that all the explicit-current-outputs are phase-shifted by 90° from each other and thus the oscillators can be used as quadrature oscillator.

3. Non-ideal Cases

For a complete analysis of the circuit, it is necessary to take into account the following CCCDTA non-ideality:

3.1 Current Tracking Errors

$$I_z = \alpha_p i_p - \alpha_n i_n \tag{18}$$

where α_p and α_n are the current transfer gains from p and n to z terminals, respectively. All these gains slightly differ from their ideal values of unity by current tracking errors of n and p input ports (ε_n and ε_p) as $\alpha_n \approx 1-\varepsilon_n$ and $\alpha_p \approx 1-\varepsilon_p$. Considering the current transfer gains, the modified characteristic equation of Figs. 3(a), (b) and (c) can be respectively expressed as

$$s^2\frac{C_1C_2R_{n1}}{(1+\alpha_{n1})\alpha_{n2}g_{m2}} + s\frac{C_2}{\alpha_{n2}g_{m2}}\left(1-\frac{\alpha_{p1}g_{m1}R_{n1}}{(1+\alpha_{n1})}\right) + \frac{\alpha_{p1}g_{m1}R_{n1}}{(1+\alpha_{n1})} = 0 , \tag{19}$$

$$s^2\frac{C_1C_2R_{n1}}{(1+\alpha_{n1})\alpha_{n2}g_{m2}} + s\frac{C_2}{\alpha_{n2}g_{m2}}\left(1-\frac{g_{m1}R_{n1}}{(1+\alpha_{n1})}\right) + \frac{g_{m1}R_{n1}}{(1+\alpha_{n1})} = 0 , \tag{20}$$

and

$$s^2\frac{C_1C_2R_{p1}}{\alpha_{n2}g_{m2}} + s\frac{C_2}{g_{m2}}\left(1-\frac{\alpha_{p1}g_{m1}R_{n1}}{(1+\alpha_{n1})}\right) + \frac{\alpha_{p1}g_{m1}R_{n1}}{(1+\alpha_{n1})} = 0 . \tag{21}$$

For non-ideal case, the oscillation condition and oscillation frequency of the proposed oscillators are as follows:

Circuit 3(a):

$$\text{OC: } 1 = \frac{\alpha_{p1}g_{m1}R_{n1}}{(1+\alpha_{n1})} , \tag{22}$$

and

$$\omega_{osc} = \sqrt{\frac{\alpha_{n2}g_{m1}g_{m2}}{C_1C_2}} . \tag{23}$$

Circuit 3(b):

$$\text{OC: } 1 = \frac{g_{m1}R_{n1}}{(1+\alpha_{n1})} , \tag{24}$$

and

$$\omega_{osc} = \sqrt{\frac{\alpha_{n2}g_{m1}g_{m2}}{C_1C_2}} . \tag{25}$$

Circuit 3(c):

$$\text{OC: } 1 = \frac{\alpha_{p1}g_{m1}R_{n1}}{(1+\alpha_{n1})} , \tag{26}$$

and

$$\omega_{osc} = \sqrt{\frac{\alpha_{n1}\alpha_{n2}g_{m1}g_{m2}}{(1+\alpha_{n1})C_1C_2}} . \tag{27}$$

It is found that parameters; α_p and α_n will affect both oscillation condition and oscillation frequency. These errors affect the sensitivity to temperature and the high frequency response of the proposed circuit. Thus, the CCCDTA should be carefully designed to minimize these errors.

3.2 Parasitic Resistances and Capacitances

The parasitic resistances and capacitances appear between the high-impedance z and x terminals of the CCCDTA and ground. The parasitic resistance and capacitances are absorbed into the external capacitance C_1 and C_2 as they appear in shunt with them. In this case, if $R_{p,n} << R_x$, R_z, the oscillation frequency for the proposed circuit in Fig. 3(a)-(b) are as follows:

Circuit 3(a):

$$\omega_{osc} = \sqrt{\frac{g_{m1}g_{m2}}{(C_1 + C_{z1})(C_2 + C_{z2})}}. \tag{28}$$

Circuit 3(b):

$$\omega_{osc} = \sqrt{\frac{g_{m1}g_{m2}}{(C_1 + C_{z1} + C_{x1} + C_{x2})(C_2 + C_{z2})}}. \tag{29}$$

Circuit 3(c):

$$\omega_{osc} = \sqrt{\frac{g_{m1}g_{m2}}{(C_1 + C_{x1} + C_{x2})(C_2 + C_{z2})}}. \tag{30}$$

To alleviate the effects of the parasitic capacitances and resistances the operating frequency ω_{osc} should be chosen such that

Circuit 3(a):

$$\omega_{osc} > \max\left[\frac{1}{(C_1 + C_{z1})R_{z1}}, \frac{1}{(C_2 + C_{z2})R_{z2}}\right]. \tag{31}$$

Circuit 3(b):

$$\omega_{osc} > \max\left[\begin{matrix}\dfrac{1}{(C_1 + C_{z1} + C_{x1} + C_{x2})R_{z1}//R_{x1}//R_{x2}}, \\ \dfrac{1}{(C_2 + C_{z2})R_{z2}}\end{matrix}\right]. \tag{32}$$

Circuit 3(c):

$$\omega_{osc} > \max\left[\frac{1}{(C_1 + C_{x1} + C_{x2})R_{x1}//R_{x2}}, \frac{1}{(C_2 + C_{z2})R_{z2}}\right]. \tag{33}$$

3.3 Nonlinearity

Nonlinearity of active devices affects the amplitude stabilization and cause both oscillation condition and oscillation frequency analyzed in (9)-(11) become aborted as well as THD becomes higher. A number of former researches have been conducted to solve this problem e.g., amplitude control by nonlinear resistors [24], [26], by AGC [24], by using inherent linearity of OTA [25], and by using a photoresistor which is a part of the 3WK16341 optron [37]. Hereby, the AGC will be added into proposed circuits. From block diagram in Fig. 2, it can be developed to Fig. 4 while AGC can be created by employing CCCDTA with a simple diode-resistor network [23] as shown in Fig. 5. From the block diagram in Fig. 4, the characteristic equation can be written as follows:

$$s^2ab + sb(1 - kk_{AGC}) + kk_{AGC} = 0. \tag{34}$$

From (34), the oscillation condition (OC) and oscillation frequency (ω_{osc}) can be expressed as

$$1 = kk_{AGC}, \tag{35}$$

and

$$\omega_{osc} = \sqrt{\frac{kk_{AGC}}{ab}}. \tag{36}$$

Although adding AGC to the circuit results in more complexity, the problem of amplitude stabilization can be finally solved and THD value becomes lower.

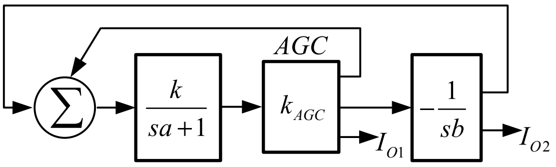


Fig. 4. Proposed circuit with AGC.

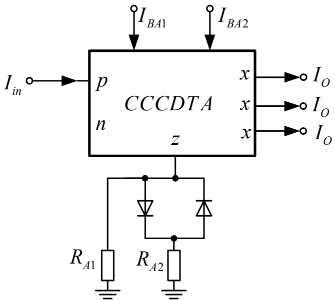


Fig. 5. CCCDTA-based AGC circuit.

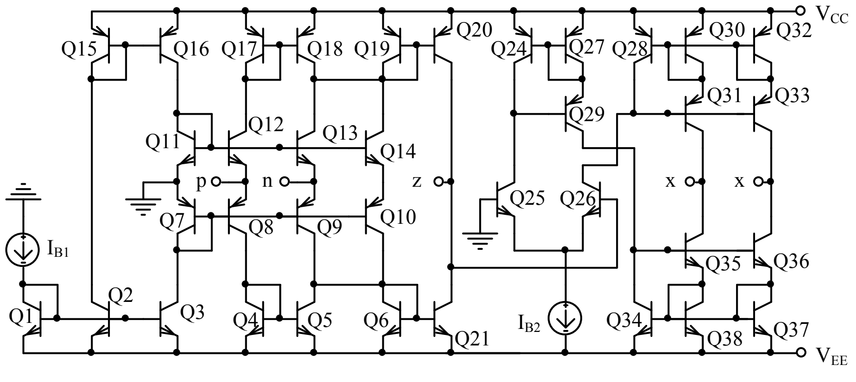


Fig. 6. Internal construction of CCCDTA.

4. Simulation Results

For example, only the proposed quadrature oscillator in Fig. 3(b) has been simulated in PSpice using the BJT implementation of the CCCDTA as shown in Fig. 6. The PNP and NPN transistors employed in the proposed circuit were simulated by using the parameters of the PR200N and NR200N bipolar transistors of ALA400 transistor array from AT&T [38]. The circuit was biased with ±2.5V supply voltages, $C_1 = C_2 = 0.4$ nF, $I_{B1} = 25$ µA, $I_{B2} = 210$ µA, $I_{B3} = 100$ µA and $I_{B4} = 180$ µA. This yields oscillation frequency of 1.23 MHz, where the calculated value of this parameter from (13) yields 1.49 MHz (deviated by 17.44%). The power consumption of the circuit is 9.25 mW. Fig. 7 shows simulated quadrature output waveforms. Fig. 8 shows the simulated output spectrum, where the total harmonic distortion (THD) is about 1.59%. The electronic tuning of the oscillation frequency with the bias current I_{B4} for different capacitor values is shown in Fig. 9.

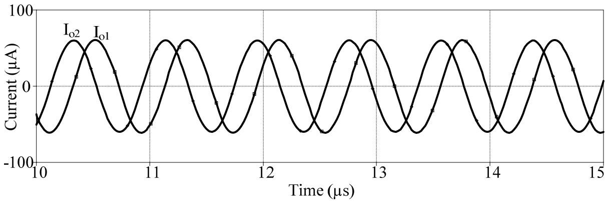


Fig. 7. Current outputs of the proposed quadrature oscillator in Fig. 3(b).

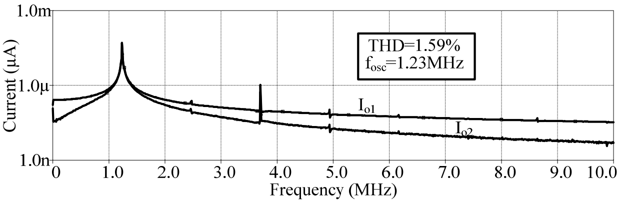


Fig. 8. Spectrum of signal in Fig.7

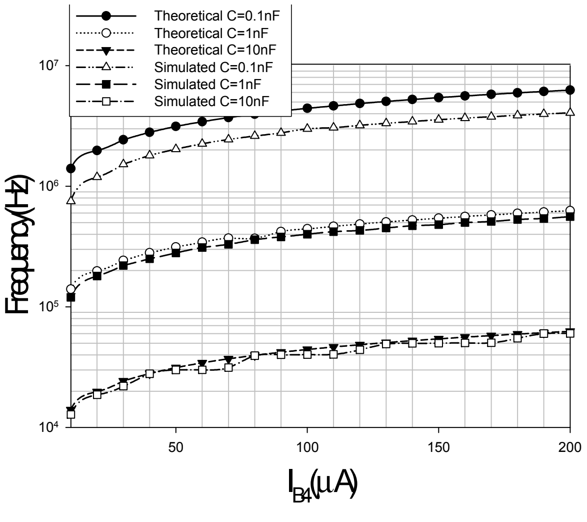


Fig. 9. Simulated oscillation frequency versus I_{B4} for different capacitances C.

5. Conclusion

The new electronically tunable current-mode quadrature oscillators based on CCCDTAs have been presented. The features of the proposed circuits are that: oscillation frequency and oscillation condition can be electronically/independently tuned; the proposed oscillators consists of merely 2 CCCCTAs and 2 grounded capacitors, non-interactive control of both the condition of oscillation and frequency of oscillation and availability of two quadrature explicit-current-outputs from high-output impedance terminals. PSpice simulation results agree well with the theoretical anticipation.

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